Adaptive Transmit Diversity with Quadrant Phase Constraining Feedback

Related Application

[01] This application is a continuation-in-part of U.S. Patent Application Sn. 10/648,558, " Adaptive Transmit Diversity with Quadrant Phase Constraining Feedback," filed on August 25, 2003 by Wu et al.

Field of the Invention

[02] This invention relates generally to transmit diversity gain in wireless communications networks, and more particularly to maximizing the diversity gain adaptively in transmitters.

Background of the Invention

[03] The next generation of wireless communication systems is required to provide high quality voice services as well as broadband data services with data rates far beyond the limitations of current wireless systems. For example, high speed downlink packet access (HSDPA), which is endorsed by the 3rd generation partnership project (3GPP) standard for wideband code-division multiple access (WCDMA) systems, is intended to provide data rates up to 10 Mbps or higher in the downlink channel as opposed to the maximum 384 Kbps supported by the enhance data rate for GSM evolution (EDGE), the so-called 2.5G communication standard, see 3GPP: 3GPP TR25.848 v4.0.0, "3GPP technical report: Physical layer aspects of ultra

high speed downlink packet access," March 2001, and ETSI. GSM 05.05, "Radio transmission and reception," ETSI EN 300 910 V8.5.1, November 2000.

[04] Antenna diversity can increase the data rate. Antenna diversity effectively combats adverse effects of multipath fading in channels by providing multiple replicas of the transmitted signal at the receiver. Due to the limited size and cost of a typical end user device, e.g., a cellular telephone or handheld computer, downlink transmissions favor transmit diversity over receiver diversity.

[05] One of the most common transmit diversity techniques is space-time coding, see Alamouti, "A simple transmit diversity technique for wireless communications," *IEEE J. Select. Area Commun.*, vol.16, pp.1451-1458, October 1998, Tarokh et al., "Space-time codes for high data rate wireless communication: performance criterion and code construction," *IEEE Trans. Info. Theory*, vol.44, pp.744-765, March 1998, Tarokh et al., "Space-time block codes from orthogonal designs," *IEEE Trans. Info. Theory*, vol.45, pp.1456-1467, July 1999, and Xin et al., "Space-time diversity systems based on linear constellation precoding," *IEEE Trans. Wireless Commun.*, vol.2, pp.294-309, March 2003.

[06] With space-time coding, data symbols are encoded in both the time domain (transmission intervals) and the space domain (transmit antenna array). For systems with exactly two transmit antennas, Alamouti et.al. describe orthogonal space-time block code (STBC). Full diversity order is achieved with simple algebraic operations.

[07] Space-time trellis coding exploits the full potential of multiple antennas by striving to maximize both the diversity gains and coding gains of the system. Better performance is achieved at the cost of relatively higher encoding and decoding complexity.

[08] The above techniques are designed under the assumption that the transmitter has no knowledge of the fading channels. Thus, those techniques can be classified as having open loop transmit diversity.

[09] System performance can be further improved when some channel information is available at the transmitter from feedback information from the receiver. Those systems are classified as having closed loop transmit diversity. The feedback information can be utilized in transmit diversity systems to maximize the gain in the receiver, see Jongren et al., "Combining beamforming and orthogonal space-time block coding," *IEEE Trans. Info. Theory*, vol.48, pp.611-627, March 2002, Zhou et al., "Optimal transmitter eigen-beamforming and space-time block coding based on channel mean feedback," *IEEE Trans. Signal Processing*, vol. 50, pp.2599-2613, October 2002, Rohani et al., "A comparison of base station transmit diversity methods for third generation cellular standards," *Porc. IEEE Veh. Techno. Conf. VTC'99 Spring*, pp.351-355, May 1999, Derryberry et al., "Transmit diversity in 3G CDMA systems," *IEEE Commun. Mag.*, vol.40, pp. 68-75, April 2002, Lo, "Maximum ratio transmission," *IEEE Trans. Commun.*, vol.47, pp. 1458-1461, October 1999, Huawe, "STTD with adaptive transmitted power allocation," TSGR1-02-0711, May, 2002, and Horng et

al., "Adaptive space-time transmit diversity for MIMO systems," *Proc. IEEE Veh. Techno. Conf. VTC'03 Spring*, pp. 1070-1073, April 2003.

[010] The space-time block coding can be combined with linear optimum beamforming. Linear encoding matrices can be optimized based on the feedback information of the fading channels. Transmit adaptive array (TxAA) is another close loop transmit diversity system with the transmitted symbols encoded only in the space domain. Increased performance can be achieved, provided the fading channel vector is known to the transmitter. The concept of space encoded transmit diversity can be generalized as maximal ratio transmission (MRT).

[011] All of the above closed loop systems require the feedback information to be $M \times N$ complex-valued matrices, where M and N are respectively the number of antennas at the transmitter and receiver. The matrix elements are either the channel impulse response (CIR), or statistics of the CIR, e.g., mean or covariance. Because the feedback matrices contain $2MN$ real-valued scalars, considerable bandwidth is consumed by the feedback information in the reverse link from the receiver to the transmitter.

[012] To overcome this problem, suboptimum methods with less feedback information are possible. Adaptive space-time block coding (ASTTD) uses a real-valued vector made up of power ratios of the fading channels as feedback information. There, the feedback information is used to adjust the power of each transmission antenna. That technique still consumes a large number of bits.

[013] Therefore, it is desired to maximize transmit diversity gain while reducing the number of bits that are fed back to the transmitter.

Summary of the Invention

[014] The invention provides an adaptive transmit diversity scheme with simple feedback for a wireless communication systems.

[015] It is an object of the invention to achieve better system performance with less feedback information and less computations than conventional transmit diversity methods.

[016] With simple linear operations at both the transmitter and receiver, the method requires only one bit of feedback information for systems with two antennas ($M = 2$) at the transmitter and one antenna at the receiver.

[017] When there are more antennas at the transmitter ($M > 2$), the number of feedback bits is $2(M - 1)$ bits. This is still significantly less than the number of bits required by most conventional closed loop transmit diversity techniques.

[018] When the indicated quadrant phase constraining method is combined with orthogonal space time block code, the amount of feedback information can be further reduced. For systems with three and four transmit antennas, the amount of feedback can be as few as one and two bits, respectively.

[019] The computational complexity of the invented method is much lower compared with optimum quantized TxAA closed loop technique with the same amount of feedback.

[020] In addition, the method outperforms some closed loop transmit diversity techniques that have more information transmitted in the feedback channel.

Brief Description of the Drawings

[021] Figure 1 is a block diagram of a system with diversity gain according to the invention;

[022] Figure 2 is a diagram of four quadrants of a coordinate system for indicating quadrant phase constraining according to the invention;

[023] Figure 3 is a diagram of a normalized coordinate system with the phase of the reference signal on the x-axis of the coordinate system according to the invention; and

[024] Figure 4 is a block diagram of a system combining orthogonal space time block code and quadrant phase constraining according to the invention.

Detailed Description of the Preferred Embodiment

[025] Figure 1 shows a baseband representation of a diversity system 100 according to our invention. Our system has M antennas 101 at a transmitter 10, for example, a base station, and one antenna 102 at a receiver 20, e.g., a cellular telephone.

[026] At a time instant k, a modulated symbol s_k 103 is linearly encoded 110 at the transmitter in a space domain according to a space encoding vector 111
$\mathbf{p}_k = [p_1(k), p_2(k), \cdots, p_M(k)] \in C^{1\times M}$.

[027] The encoded transmit data 112 are $\mathbf{x}_k = [x_1(k), x_2(k), \cdots, x_M(k)] = \mathbf{p}_k \cdot s_k$, with $x_m(k)$ being transmitted at the m^{th} transmit antenna 101.

[028] In our adaptive transmit diversity method, the space encoding vector $\mathbf{p}_k$ 111 is determined 120 at the transmitter according to feedback information 121 determined from space decoding 130 of the received signal 105 at the receiver.

[029] Specifically, the feedback information 121 relates to phase differences between pairs of received signals in a fading transmission channel 115. It is desired to minimize the phase difference between signals, so that diversity gain is maximized at the receiver. Furthermore, it is desired to minimize the number of bits required to indicate the phase difference. It is also desired to reduce the amount of computation involved generating the feedback information at the receiver 20.

[030] The received signal is a sum of the propagation signals from all the transmit antennas subject to the channel impulse responses, plus additive white Gaussian noise (AWGN) 104 with variance $N_0/2$ per dimension. At the receiver, samples r_k 113 of the received signal R_x can be expressed by

$$
\begin{aligned}
r(k) &= \sqrt{\frac{E_s}{M}} \cdot \mathbf{x}_k \mathbf{h}_k + z_k, \\
&= \sqrt{\frac{E_s}{M}} \cdot (\mathbf{p}_k \mathbf{h}_k) \cdot s_k + z_k,
\end{aligned} \qquad (1)
$$

where E_s is the sum of the transmit energy of all the transmit antennas, M is the number of antennas, z_k is the additive noise 104. The time-varying channel impulse response (CIR) of each fading channel is $\mathbf{h}_k = [h_1(k), h_2(k), \cdots, h_M(k)]^T \in C^{M \times 1}$, where $h_m(k)$ is the CIR for the fading channel between the m^{th} transmit antenna and the receive antenna, and $(\cdot)^T$ denotes a matrix transpose.

[031] With the system model defined by equation (1), an optimum space encoding vector $\hat{\mathbf{p}}_k$ for maximizing the output SNR is

$$
\hat{\mathbf{p}}_k = \frac{\mathbf{h}_k^H}{\mathbf{h}_k \mathbf{h}_k^H}, \qquad (2)
$$

where $(\cdot)^H$ denotes a Hermitian matrix operator. This scheme is called transmit adaptive array (TxAA). However, forming the optimum space encoding vector requires a complete of understanding of the CIR vector $\mathbf{h}_k$, which contains $2M$ real scalar values. Hence, it is impractical to implement the TxAA scheme in practical systems where limited resources are allocated to the feedback channel.

[032] To reduce the amount of feedback information, an optimum quantized feedback scheme is described for TxAA. The space encoding vector is obtained from an exhaustive searching algorithm as follows

$$\hat{\mathbf{p}}_k = \arg\min_{\mathbf{p}_k \in P} \mathbf{p}_k \mathbf{h}_k \mathbf{h}_k^H \mathbf{p}_k^H \tag{3}$$

where P is the set of all the possible quantized space encoding vectors. The set contains $2^{b(M-1)}$ possible vectors for systems with b bits quantization and M transmit antennas. In order to find the optimum quantized feedback vector $\hat{\mathbf{p}}_k$, the receiver must exhaustively determine the values of $\mathbf{p}_k \mathbf{h}_k \mathbf{h}_k^H \mathbf{p}_k^H$ for all the possible $2^{b(M-1)}$ encoding vectors before the optimum encoding vector can be selected.

[033] Each computation of the cost function involves approximately M^2 complex multiplications. Therefore, the total amount of computational complexity incurred by the feedback information alone is in the order of $O(2^{b(M-1)} \times M^2)$, which increases exponentially with the number of transmit antennas and is quite considerable when the number of antennas is larger than two.

[034] To balance the system performance, the size of feedback information, and the computational complexity of the system, the adaptive transmit diversity method according to our invention uses a quadrant phase constraining method to determine the feedback information. Thus, both the amount of feedback and computation complexity can be greatly reduced.

[035] Method Description

[036] The present adaptive transmit diversity method is described first for the simplest system with two transmit antennas and one receive antenna. In this simple case, exactly *one* bit of feedback information is required to generate the space encoding vector 111 used by the space encoding 110. In a general method for systems with $M > 2$ transmit antennas, it takes $2(M - 1)$ bits of feedback information to determine 120 the space encoding vector 111.

[037] Systems with Two Transmit Antennas

[038] For systems with two transmit antennas, we define our space encoding vector 111 as

$$\mathbf{p}_k = [1, (-1)^{b_k}], \tag{4}$$

where $b_k \in \{0,1\}$ is the quantized binary feedback information 121 sent out from the receiver. The single feedback bit b_k, either zero or one, is based on an estimated phase shift in the CIR $\mathbf{h}_k$ as follows,

$$b_k = \begin{cases} 0, & \text{if } \Re\left\{h_1(k)h_2^*(k)\right\} > 0, \\ 1, & \text{otherwise}, \end{cases} \tag{5}$$

where $h_m(k)$ is the time-varying channel impulse response, $(\cdot)^*$ denotes a complex conjugate, and the operation $\Re(\cdot)$ returns the real part of the operand. In other words, the bit is zero if the product of the CIR of one channel with the complex conjugate of the CIR of the other channel is positive, and one otherwise, and thus, the space encoding vector $\mathbf{p}$ 111 is either [1,1] or [1, -1], respectively.

[039] With the definition of the space encoding vector $\mathbf{p_k}$ in equation (3), the transmitted signal vector 112 is $\mathbf{x}_k = [s_k, (-1)^{b_k} s_k]$. Replacing the vector $\mathbf{x}_k$ in equation (1), we have the received sample as

$$r(k) = \sqrt{\frac{E_s}{2}}[h_1(k) + h_2(k)(-1)^{b_k}] \cdot s_k + z_k. \tag{6}$$

[040] In receivers with coherent detection, the received sample $r(k)$ is multiplied by $(\mathbf{p}_k \mathbf{h}_k)^H = h_1^*(k) + h_2^*(k)(-1)^{b_k}$ to form the decision variable $y(k)$,

$$y(k) = (h_1^*(k) + (-1)^{b_k} h_2^*(k)) \cdot r_k,$$

$$= \sqrt{\frac{E_s}{2}}\left[|h_1(k)|^2 + |h_2(k)|^2 + (-1)^{b_k} \cdot 2\Re\{h_1(k)h_2^*(k)\}\right] s_k + v_k, \tag{7}$$

where $v_k = \left[h_1^*(k) + (-1)^{b_k} h_2^*(k)\right] \cdot z_k$ is the noise component of the decision variable. The variance of noise component v_k is

$$\sigma_v^2 = \left[|h_1(k)|^2 + |h_2(k)|^2 + (-1)^{b_k} \cdot 2\Re\{h_1(k)h_2^*(k)\}\right] \cdot N_0. \tag{8}$$

[041] It can be seen from equation (5) that

$$(-1)^{b_k} \cdot 2\Re\{h_1(k)h_2^*(k)\} = 2|\Re\{h_1(k)h_2^*(k)\}|, \tag{9}$$

thus the instantaneous output SNR γ at the receiver can be written as

$$\gamma = \frac{\gamma_0}{2} \cdot \left[|h_1(k)|^2 + |h_2(k)|^2 + 2|\Re\{h_1(k)h_2^*(k)\}|\right], \tag{10}$$

$$= \gamma_0 \cdot (g_c + g_b), \tag{11}$$

where $\gamma_0 = \frac{E_s}{N_0}$ is the SNR without diversity. The conventional diversity gain g_c and the feedback diversity gain g_b are defined as

$$g_c = \frac{1}{2}[|h_1(k)|^2 + |h_2(k)|^2], \tag{12}$$

$$g_b = 2\,|\,\Re\{h_1(k)h_2^*(k)\}\,|\,. \qquad (13)$$

The conventional diversity gain g_c is the same as the diversity gain of the orthogonal space-time block coding (STBC), while the feedback diversity gain g_b is the extra diversity gain contributed by the binary feedback information 121.

[042] From the above equations, we can see that with only one bit b_k of feedback information 121 in a closed loop system, the output SNR of our transmission diversity scheme, which also considers feedback diversity gain, is always better than when just the orthogonal STBC gain is considered in an open loop system, although the transmitted signals are only encoded in the space domain.

[043] **Systems with More Than Two Transmit Antennas**

[044] The process described above is for systems with two transmit antennas. If there are more than two antennas ($M > 2$) at the transmitter, then a modified transmit diversity method with $2(M - 1)$ bits feedback information is used.

[045] For systems with $m > 2$ transmit antennas, we define the space encoding vector 111 as

$$\mathbf{p}_k = \left[1 \quad \exp[\frac{i \cdot q_2(k)\pi}{2}] \quad \cdots \quad \exp[\frac{i \cdot q_M(k)\pi}{2}]\right], \qquad (14)$$

where $i^2 = -1$, and $q_m(k) \in \{0,1,2,3\}$ is the feedback information from the receiver, for $m = 2,3,\cdots,M$. For consistence of representation, we let $q_1(k) = 0$, for $\forall k$.

[046] By such definitions, each $q_m(k)$ contains two bits of information, and there are a total of $2(M - 1)$ bits of feedback information used to form the space encoding vector $\mathbf{p}_k$. Combining equations (1) and (14), we can write the received sample $r(k)$ as

$$r(k) = \sqrt{\frac{E_s}{M}} \left\{ \sum_{m=1}^{M} \exp\left[\frac{i \cdot q_m(k)\pi}{2} \right] h_m(k) \right\} \cdot s_k + z_k. \tag{15}$$

[047] At the decoder 130, the decision variable $y(k)$ is obtained by multiplying the received sample $r(k)$ with $(\mathbf{p}_k \mathbf{h}_k)^H$. This can be written as

$$\begin{aligned} y(k) &= \sqrt{\frac{E_s}{M}} \left| \sum_{m=1}^{M} \exp[\frac{i \cdot q_m(k)\pi}{2}] h_m(k) \right|^2 \cdot s_k + v_k, \\ &= \sqrt{\frac{E_s}{M}} (g_c + g_b) \cdot s_k + v_k, \end{aligned} \tag{16}$$

where $v_k = (\mathbf{p}_k \mathbf{h}_k)^H \cdot z_k$ is the noise component with variance $|\mathbf{p}_k \mathbf{h}_k|^2 \cdot N_0$, and the conventional and feedback diversity gains g_c and g_b are defined respectively as

$$g_c = \frac{1}{M} \sum_{m=1}^{M} |h_m(k)|^2, \text{ and} \tag{17}$$

$$g_b = \frac{2}{M} \sum_{m=1}^{M} \sum_{n=m+1}^{M} \Re\left\{ h_m(k) h_n^*(k) \exp\left[i \cdot \pi \frac{q_m(k) - q_n(k)}{2} \right] \right\}. \tag{18}$$

[048] In the equations above, the conventional diversity gain g_c is fixed for a certain value of M, while the feedback diversity gain is maximized by appropriately selecting the feedback information based on $q_m(k)$.

[049] With the 2(M - 1) bits of information, we can maximize g_b by selecting $q_m(k)$ such that all the summed elements of g_b are positive. One of the summed element of g_b can be expressed as

$$\Re\left\{h_m(k)h_n^*(k)\exp\left[i\cdot\pi\frac{q_m(k)-q_n(k)}{2}\right]\right\}=|h_m(k)||h_n(k)|\cos(\Delta\theta_{mn}),\quad (19)$$

where $\theta_m\in[0,2\pi)$ is the phase of $h_m(k)$.

[050] The terms in equation (19) are positive when the following condition is satisfied

$$|\Delta\theta_{mn}|\le\pi/2,\text{for }\forall m\ne n.\qquad (20)$$

In words, the absolute difference in phase between two signals is less than 90 degrees.

[051] To satisfy this maximization condition in equation (20), we adjust $q_m(k)$ so that the phases $\theta_m+\frac{q_m(k)}{2}\pi$, for $m=1,2,\cdots,M$, for all received signals are within 90 degrees of each other. We call this method a quadrant phase constraining method.

[052] Without loss of generality, we keep the phase θ_1 of the signal in the first sub-channel $h_1(k)$ unchanged. We call this the reference phase of the reference signal. The reference phase can be selected arbitrarily from any of the M transmit antennas, or the CIR with the highest power.

[053] Now, the goal is to make the phases difference between all the signals less than $\frac{\pi}{2}$, or constraining all the shifted phases to a quadrant phase sector, i.e., a sector of 90 degrees.

[054] Therefore, the phases θ_m of the signals in all other sub-channels need to be rotated counter-clockwise at the transmitter $\frac{q_m(k)}{2}\pi$ so that the absolute phase difference is less than 90 degrees. By such means, only two bits of information are required to form each $q_m(k)$.

[055] One method to fulfill the quadrant phase constraining condition is to put all the phases in the same coordinate quadrant as the reference phase. As shown in Figure 2, we label four quadrants I-IV of the Cartesian coordinate system for real (Re) and imaginary (Im) numbers. The quadrant number of any angle $\phi \in [0, 2\pi)$ is $\left\lceil \frac{2\phi}{\pi} \right\rceil$, where $\lceil \square \rceil$ denotes rounding up to the nearest integer.

[056] With the above analyses, the feedback information $q_m(k)$ for $m = 2, 3, \cdots, M$ is determined at the receiver based on the phase difference between any pair of received signals.

$$q_m(k) = \left\lceil \frac{2\theta_1}{\pi} \right\rceil - \left\lceil \frac{2\theta_m}{\pi} \right\rceil. \qquad (21)$$

[057] The example 200 in Figure 2 has θ_1 in quadrant II, and θ_m in quadrant IV. With equation (21), we obtain $q_m(k) = -2$, which corresponds to

rotate θ_m by π radians clockwise (180°), and the rotated phase $\theta_m - \frac{q_m(k)}{2}\pi$ is now in quadrant II.

[058] Alternatively, all the phases are put in a 90 degree sector 300 centered around the reference phase as shown in Figure 3. We normalize all the phases with respect to the reference phase as follows $\tilde{\theta}_m = \theta_m - \theta_1 + 2l\pi$, where the integer l is chosen such that the normalized phase $\tilde{\theta}_m$ is in the range of $[0, 2\pi)$. The normalized phase $\tilde{\theta}_m$ is rotated counter-clockwise by the angle of $q_m\frac{\pi}{2}$, so that the rotated angle $\tilde{\theta}_m + q_m\frac{\pi}{2}$ is in the quadrant phase sector from $[-\pi/4, \pi/4]$ of the coordinate system as shown in Fig. 3. Following the description above, we can compute the feedback information q_m as

$$q_m = \begin{cases} 4 - \left\lfloor \dfrac{\tilde{\theta}_m + \pi/4}{\pi/2} \right\rfloor, & \tilde{\theta}_m \in [\dfrac{\pi}{4}, \dfrac{7\pi}{4}), \\ 0, & \text{otherwise}, \end{cases}$$

where $\lfloor \square \rfloor$ returns the nearest smaller integer. An example is shown in Figure 3, where $\tilde{\theta}_m = 9\pi/8$.

[059] We can determine that $q_m = 2$, and the corresponding rotated angle is $\tilde{\theta}_m - q_m\frac{\pi}{2} = \pi/8$, which is in the quadrant phase sector of $[-\frac{\pi}{4}, \frac{\pi}{4}]$ of the coordinate system.

[060] By performing the same operations one by one to all of the normalized phases, the rotated phases are confined to the same quadrant phase sector, and the non-negativity of each summed element of the diversity gain g_b can be guaranteed.

[061] This method achieves the non-negativity of the feedback diversity gained by constraining all the rotated phases of the CIRs of one group of transmit antennas in a quadrant phase sector of $\pi/2$. Hence, we call it quadrant phase constraining method.

[062] Because the feedback value of q_m is determined independently for each of the transmit antennas, the computational complexity of our method increases linearly with the number of transmit antennas, as opposed to the exponentially increased complexity of the prior art optimum quantization method.

[063] The feedback information computed from the quadrant phase constraining method guarantees that all the elements described in Equation (19) are positive for $\forall m \neq n$, and the maximized feedback diversity gain g_b contributed by the feedback information is written as

$$g_b = \frac{2}{M}\sum_{m=1}^{M}\sum_{n=m+1}^{M} |h_m(k)||h_n(k)||\cos(\Delta\theta_{mn})|. \qquad (22)$$

Combining equations (16), (17) and (22), yields the output SNR γ at the detector receiver as

$$\gamma = \gamma_0 \cdot \left[\frac{1}{M}\sum_{m=1}^{M}|h_m(k)|^2 + \frac{2}{M}\sum_{m=1}^{M}\sum_{n=m+1}^{M}|h_m(k)||h_n(k)||\cos(\Delta\theta_{mn})|\right], \qquad (23)$$

$$= \gamma_0 \cdot (g_c + g_b), \qquad (24)$$

where $\gamma_0 = \frac{E_s}{N_0}$ is the SNR without diversity, and the diversity gains g_c and g_b are given in equations (17) and (22), respectively.

[064] Complexity Analysis

[065] As described above, the method according to the invention determines feedback information for each transmit antenna separately. Therefore, the computation complexity increases only linearly with the number of transmit antennas. However, for the optimum quantized feedback TxAA method, the computational complexity increases exponentially with the number of transmit antennas. For systems with $M = 4$ transmit antennas and two bits representation of each element of the space encoding vector $\mathbf{p}_k$, there are totally $2^{2\times(4-1)} = 64$ possible values of $\mathbf{p}_k$. This means that a receiver employing optimum quantized feedback must compute $\mathbf{p}_k \mathbf{h}_k \mathbf{h}_k^H \mathbf{p}_k^H$ for all the 64 possible vectors of $\mathbf{p}_k$ before the feedback information can be sent, and each computation of the cost function $\mathbf{p}_k \mathbf{h}_k \mathbf{h}_k^H \mathbf{p}_k^H$ involves approximately $4^2 = 16$ COMPLEX multiplications. However, our sub-optimum method requires only M - 1 = 3 computations for all the antennas, and each operation involves approximately 2 REAL multiplications. Therefore, the computational complexity of our method is only $(2\times 3)/(64\times 16\times 2) = 0.3\%$ of the prior art optimum quantized feedback TxAA for system with M = 4 transmit antennas. For system with more transmit antennas, even larger computational complexity saving can be achieved by our method.

[066] Combining Orthogonal STBC with Group Space Encoding

[067] The method described above only involves the encoding process in the space domain. To further reduce the amount of feedback and

computation, the quadrant phase constraining feedback scheme is combined with orthogonal space-time block coding (STBC). In this method, the time domain is also utilized in the encoding process.

[068] The system structure is shown in Figure 4. Input symbols 401 are generated and modulated by conventional means. The symbols are fed into an orthogonal STBC encoder 410. Without loss of generality, we assume that at two consecutive symbol periods t_1 and t_2, an input to the STBC encoder is s_1 and s_2, respectively, where $s_j \in S$, for $j = 1,2$, with S being the modulation symbol set.

[069] The energy of the modulation symbol is $E(|s_j|^2) = E_s$. At the STBC encoder, the input data symbols s_1 and s_2 are demultiplexed into multiple data streams, one for each group of transmit antennas. The data stream 411 of the STBC encoder 410 is expressed by

$$\begin{aligned} \mathbf{d}_1 &= [d_{11} \quad d_{21}]^T = [s_1 \quad s_2^*]^T \in C^{2\times 1}, \\ \mathbf{d}_2 &= [d_{12} \quad d_{22}]^T = [s_2 \quad -s_1^*]^T \in C^{2\times 1}, \end{aligned} \tag{25}$$

where d_k corresponds to the k^{th} output stream of the STBC encoder, with d_{kj} being transmitted at the time instant t_j, and $(\cdot)^T$ denotes matrix transpose.

[070] The M transmit antennas are divided into multiple groups of transmit antennas 421-422. Each group corresponds to one of the data streams $\mathbf{d}_1$, $\mathbf{d}_2$ produced by the STBC encoder 410. We assume the number of antennas contained in the kth group is M_k, for $k = 1,2$, with $M_1 + M_2 = M$.

[071] Adaptive linear space encoders 431-432 are applied to each data stream 411 for each group of transmit antennas. The space encoders 431-432 map the multiple data streams 411 to the groups of transmit antennas according to channel feedback information 440 for each group.

[072] If we define a space encoding vector of the k^{th} group as

$$\mathbf{p}_k = [p_{k,1} \quad p_{k,2} \quad \cdots \quad p_{k,M_k}] \in C^{1\times M_k}, \text{for } k = 1, 2, \tag{26}$$

with the constraint $\mathbf{p}_1\mathbf{p}_1^H + \mathbf{p}_2\mathbf{p}_2^H = 1$, then encoded signals 433 to be transmitted by the k^{th} antenna group can be expressed in matrix format

$$\mathbf{X}_k = \mathbf{d}_k \cdot \mathbf{p}_k \in C^{2\times M_k}, \text{for } k = 1, 2, \tag{27}$$

with the symbols on the first row of $\mathbf{X}_i$ transmitted at the symbol period t_1 and symbols on the second row transmitted at t_2.

[073] In the channel, the received signals 461 are corrupted by both time-varying mulitpath fading and AWGN 462.

[074] A receiver 450 includes a space-time decoder 451, a channel estimation module 452, and a feedback computation unit 453 for generating the feedback information 440 for each group of transmit antennas. The signals *Rx* 461 received by the receiver 450 are the sum of the propogational signals from all the transmit antennas plus the noise 462. The received signals can be represented by

$$\begin{aligned} \mathbf{r} &= [\mathbf{X}_1 \quad \mathbf{X}_2]\begin{bmatrix}\mathbf{h}_1 \\ \mathbf{h}_2\end{bmatrix} + \mathbf{z}, \\ &= \mathbf{d}_1\mathbf{p}_1\mathbf{h}_1 + \mathbf{d}_2\mathbf{p}_2\mathbf{h}_2 + \mathbf{z}, \end{aligned} \tag{28}$$

where $\mathbf{r} = [r_1, r_2]^T$, $\mathbf{z} = [z_1, z_2]^T$ are the receive vector and AWGN noise vector,

respectively, with r_k and z_k corresponding to the time instant t_k, $\mathbf{h}_k \in C^{M_k \times 1}$ is the channel impulse response (CIR) defined as

$$\mathbf{h}_k = \begin{bmatrix} h_{k,1} & h_{k,2} & \cdots & h_{k,M_k} \end{bmatrix}^T, \text{ for } k = 1, 2, \tag{29}$$

with the element $h_{k,m}$, for $m = 1, 2, \cdots, M_k$, being the CIR between the m^{th} transmit antenna of group k and the receive antenna.

[075] Combining Equations (1) and (5), we can rewrite the input-output relationship of the diversity system as

$$\begin{bmatrix} r_1 \\ r_2^* \end{bmatrix} = \begin{bmatrix} \mathbf{p}_1\mathbf{h}_1 & \mathbf{p}_2\mathbf{h}_2 \\ -\mathbf{h}_2^H\mathbf{p}_1^H & \mathbf{h}_1^H\mathbf{p}_1^H \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} + \begin{bmatrix} z_1 \\ z_2^* \end{bmatrix},$$

$$= \mathbf{H} \cdot \mathbf{s} + \mathbf{z}, \tag{30}$$

where $(\cdot)^*$ denotes complex conjugate, $\mathbf{s} = [s_1\ s_2]^T$ is the signal vector, and the channel matrix $\mathbf{H}$ is defined as

$$\mathbf{H} = \begin{bmatrix} \mathbf{p}_1\mathbf{h}_1 & \mathbf{p}_2\mathbf{h}_2(12) \\ -\mathbf{h}_2^H\mathbf{p}_1^H & \mathbf{h}_1^H\mathbf{p}_1^H \end{bmatrix} \in C^{2\times 2}. \tag{31}$$

The matrix $\mathbf{H}$ is an 2×2 orthogonal matrix, i.e., $\mathbf{H}^H\mathbf{H} = (|\mathbf{h}_1\mathbf{w}_1|^2 + |\mathbf{h}_2\mathbf{w}_2|^2) \cdot \mathbf{I}_2$, with $\mathbf{I}_2$ being a 2×2 identity matrix. From Equations (11) and (13), we can determine the decision vector $\mathbf{y} = [y_1, y_2]^T$ as $\mathbf{y} = \mathbf{H}^H\mathbf{r}$,

$$= (|\mathbf{h}_1\mathbf{p}_1|^2 + |\mathbf{h}_2\mathbf{p}_2|^2) \cdot \mathbf{s} + \mathbf{v}, \quad (32)$$

where $\mathbf{v} = \mathbf{H}^H\mathbf{z}$ is the noise component with covariance matrix

i. $(|\mathbf{h}_1\mathbf{p}_1|^2 + |\mathbf{h}_2\mathbf{p}_2|^2)\mathbf{I}_2 \cdot N_0$, and $N_0 = E(|z_k|^2)$.

[076] With the decision variable given in Equation (14), we can compute the signal to noise ratio at the receiver as follows

$$\gamma = (|\mathbf{h}_1\mathbf{p}_1|^2 + |\mathbf{h}_2\mathbf{p}_2|^2)\cdot\gamma_0, \tag{33}$$

where $\gamma_0 = \frac{E_s}{N_0}$ is the SNR without diversity. It can be seen from Equation (15) that the SNR γ is a function of the space encoding vectors $\mathbf{p}_1$, $\mathbf{p}_2$ and the CIR vectors $\mathbf{h}_1$, $\mathbf{h}_2$.

[077] By selecting appropriate forms of $\mathbf{p}_k$, based on the properties of the fading channels, we can improve the receiver SNR with only a small amount of feedback information 440.

[078] In our method, we apply the quadrant phase constraining feedback method in the design of the group space encoding vector $\mathbf{p}_k$ to save both the computational complexity and feedback amount. These details are described below.

[079] Space Encoding Vector Design: General Case

[080] To achieve the maximum SNR at the receiver, the optimum design criterion for the space encoding vectors $\mathbf{w}_1$ and $\mathbf{w}_2$ is

$$(\mathbf{p}_1, \mathbf{p}_2) = \underset{(\mathbf{p}_1,\mathbf{p}_2)\in W}{\arg\max} |\mathbf{h}_1\mathbf{p}_1|^2 + |\mathbf{h}_2 p_2|^2, \tag{34}$$

where W is the set of all the possible encoding vector pairs with the constraint $\mathbf{p}_1\mathbf{p}_1^H + \mathbf{p}_2\mathbf{p}_2^H = 1$. The optimum values of $\mathbf{p}_1$ and $\mathbf{p}_2$ can be obtained by exhaustive search of all the elements of W. The size of the set W increases exponentially with the number of transmit antennas, therefore this optimum space encoding vector design method is inappropriate for systems with large number of transmit antennas.

[081] In order to reduce the computational complexity, as well as to reduce the amount of feedback information, we apply the quadrant phase constraining method for the computation of the feedback information and the formulation of the adaptive space encoding vectors.

[082] For a general system with M transmit antennas, we let $M_1 = M_2 = \frac{M}{2}$ when M is an even number, and $M_1 = \frac{M+1}{2}$, $M_2 = \frac{M-1}{2}$ when M is an odd number. We define the space encoding vector $\mathbf{p}_k$ as

$$\mathbf{p}_k = \frac{1}{\sqrt{M}}\left[1 \quad \exp\left(-\mathbf{i}\cdot q_{k,2}\frac{\pi}{2}\right) \quad \cdots \quad \exp\left(-\mathbf{i}\cdot q_{k,M_k}\frac{\pi}{2}\right)\right] \in C^{1\times M_k}, \text{for } k = 1,2, \tag{35}$$

where $\mathbf{i}^2 = -1$ is the imaginary part symbol, $q_{k,m} \in \{0,1,2,3\}$, for $m = 2,3,\cdots,M_k$ and $k = 1,2$, is the feedback information, and each $q_{k,m}$ contains two bits of information. For systems with M transmit antennas, the total number of feedback bits required by our method is $2M-4$. For convenience of representation, we let $q_{1,1} = q_{2,1} = 0$.

[083] Applying the quadrant phase constraining method, we can compute the feedback information $q_{k,m}$ as

$$q_{k,m} = \begin{cases} \left\lfloor \dfrac{\tilde{\theta}_{m,k} + \pi/4}{\pi/2} \right\rfloor, & \tilde{\theta}_{k,m} \in [\frac{\pi}{4}, \frac{7\pi}{4}), (24) \\ 0, & \text{otherwise}, \end{cases} \tag{36}$$

where $\lfloor \square \rfloor$ returns the nearest smaller integer, and

$$\tilde{\theta}_{k,m} = \theta_{k,m} - \theta_{k,1} + 2l\pi, \tag{37}$$

with the integer l selected such that $\tilde{\theta}_{k,m}$ is in the range of $[0, 2\pi)$.

[084] With the adaptive diversity algorithm described here, $2M-4$ bits of feedback information are required to form the space encoding vectors for systems with M transmit antennas. It will be shown next that the amount of feedback information can be further reduced for systems with $M=4$ or $M=3$ transmit antennas, which are of practical interests of next generation communication systems.

[085] Space Encoding Vector Design: Special Case

[086] For systems with $M \le 4$ transmit antennas, each group has two transmit antennas at most. For groups with two transmit antennas, our sub-optimum design criterion can be satisfied with only one bit of feedback information.

[087] For a systems with $M=4$ transmit antennas, the number of antennas in each of the antenna groups is $M_1 = M_2 = 2$. We define the space encoding vector as

$$\mathbf{p}_k = \frac{1}{2}[1 \ (-1)^{b_k}], \text{ for } k=1,2, \qquad (38)$$

where $b_k \in \{0,1\}$ is the feedback information for the k^{th} antenna group. The feedback information can be defined by

$$b_k = \begin{cases} 0, & \Re(h_{k,1} h^*_{k,2}) \ge 0, \\ 1, & \text{otherwise} \end{cases} \qquad (39)$$

[088] The SNR at the receiver is expressed by

$$\gamma_4 = (g_{4,c} + g_{4,b})\gamma_0, \qquad (40)$$

with the conventional diversity gain $g_{4,c}$ and the feedback diversity gain $g_{4,b}$

defined as

$$g_{4,c} = \frac{1}{4}\left(\sum_{m=1}^{2} |h_{1,m}|^2 + \sum_{m=1}^{2} |h_{2,m}|^2\right), \qquad (41)$$

$$g_{4,b} = \frac{1}{2}\sum_{k=1}^{2} |\Re(h_{k,1} h_{k,2}^*)|. \qquad (42)$$

[089] Similarly, for systems with $M = 3$ antennas, we have groups $M_1 = 2$ and $M_2 = 1$. Because there is only one antenna in the second group, we have $\mathbf{p}_2 = 1/\sqrt{3}$.

[090] For the first group with two transmit antennas, we apply the space encoding vector $\mathbf{p}_1$. With this encoding scheme, the receiver SNR can be computed from Equation (15) as

$$\gamma_3 = (g_{3,c} + g_{3,b}) \cdot \gamma_0, \qquad (43)$$

with the conventional diversity gain $g_{3,c}$ and feedback diversity gain $g_{3,b}$ given by

$$g_{3,c} = \frac{1}{3}\left(\sum_{m=1}^{2} |h_{1,m}|^2 + |h_{2,1}|^2\right), \qquad (44)$$

$$g_{3,b} = \frac{2}{3} |\Re(h_{k,1} h_{k,2}^*)|. \qquad (45)$$

[091] When there are only two transmit antennas in the system, we have $w_1 = w_2 = 1/\sqrt{2}$, and this scheme is reduced to orthogonal space time block coding described above.

[092] With our method, we only need one bit and two bits of feedback information for systems with $M = 3$ and $M = 4$ transmit antennas, respectively.

[093] Performance Bounds

[094] Based on the statistical properties of the output signal 105 at the receiver 20, the theoretical performance bounds of our diversity scheme as

$$P^{U}(E) = \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \prod_{m=1}^{M}\left(1+\frac{\bar{\gamma}_m}{\sin^2\theta}\right)^{-1} d\theta, \tag{46}$$

$$P^{L}(E) = \frac{1}{2} - \frac{1}{\pi}\int_0^{\pi/2} \frac{\exp(-\tan\theta)}{\sin 2\theta} \Im\left\{\phi(2\sqrt{\tan\theta})\right\} d\theta. \tag{47}$$

Here, the derivations of $P^{U}(E)$ and $P^{L}(E)$ are omitted for the purpose of clarity. With the theoretical performance bounds given in equation (46) and (47), the actual error probability $P(E)$ of our diversity scheme satisfies

$$P^{U}(E) \geq P(E) \geq P^{L}(E). \tag{48}$$

[095] Equations (46 - 48) evaluate the method according to the invention on a theoretical basis, and these equations can be used as a guide for designing wireless communication systems.

[096] It should be noted that the conventional full-rate STBC and close loop technique based on the orthogonal STBC can only be implemented for systems with exactly two transmit antennas.

[097] In contrast, the transmit diversity method according to the invention can be used for systems with an arbitrary number of transmit antennas. This is extremely useful for a high speed downlink data transmission of next generation wireless communication systems, where higher diversity orders are required to guarantee high data throughput in the downlink with multiple transmit antennas and one receive antenna.

Effect of the Invention

[098] The method according to the invention outperforms conventional orthogonal STBC by up to 2 dB. The performance of the version with two bits of feedback information is approximately 0.4 dB better than the version with one bit of feedback information.

[099] The prior art full rate STTD and ASTTD systems can be implemented for systems with at most two transmit antennas. In contrast, our transmit diversity method can be used for systems with an arbitrary number of transmit antennas. Furthermore, the performance of the method improves substantially linearly with the increasing number of transmit antennas.

[0100] Our method is very computationally efficient compared to the prior art optimum quantized method. Our method requires only 0.3% computation efforts of the prior art optimum quantized feedback TxAA for systems with 4 transmit antennas. This computation saving is significant at the receiver, which is usually a battery powered cellular phone.

[0101] Although the invention has been described by way of examples of preferred embodiments, it is to be understood that various other adaptations and modifications can be made within the spirit and scope of the invention. Therefore, it is the object of the appended claims to cover all such variations and modifications as come within the true spirit and scope of the invention.